A Method of Broad-Band Matching to SAW-IDT (surface acoustic wave interdigital transducer)

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A method is described for designing a broad-band matching network between IDT and electrical source or load, by use of only one transformer and one tuning inductor. The method is based upon both detuning the resonance frequency of the matching network and decreasing the network Q factor by the step-up transformer. Calculations show feasibility of SAW delay line having insertion loss (IL) of 21 dB, 3 dB fractional bandwidth (BW) of 55% and triple transit suppression (TTS) of 62 dB for 3 electrode (N=1) IDT. An experimental example shows IL=25 dB, BW=46.5% and TTS=61 dB around at 33 MHz, and confirms the theoretical predictions except for BW. The short IDT launchs spurious bulk modes and bulk longitudinal wave degrades substantially BW.

I Introduction

Increase of efficiency and bandwidth of interdigital transducers (IDTs) results in remarkable improvement on the performance of wide-band surface acoustic wave (SAW) devices such as SAW delay lines and more sophistcated signal processing devices,[1].

Since the IDT has its own frequency characteristics and is mostly electrically reactive, there is a constraint upon the impedance matching between electrical source (or load) and IDT to be maintained over a wide-band range. In order to overcome this problem, there have been numerous studies on IDT structure with controlling interdigital spacing and overlap (apodization) to achieve the desired frequency response[2]. By this method even octave bandwidth can be obtained, but untuned operation of the devices increases transducer insertion loss and a large number of electrodes i. e., large acoustic length, are required to obtain a smooth frequency response, increasing the fabrication difficulty. Moreover, in spite of sophistcated design, these electrodes give rise to amplitude ripple, especially at the band edges[2]. On the other hand other workers demonstrated a broad-band coupling method for the conventional uniform IDT with short acoustic length, by use of lumped inverter networks[3~5]. They obtained 55% 3 dB-fractional bandwidth centered at 235 MHz with 25 dB insertion loss and 50 dB triple-transit suppression from a pair of 3-electrode IDTs[4]. Since the inverter network is approximated by lumped element π sections, physical size of inductors becomes large and loss due to these elements can not be ignored for operation at frequency lower than 100 MHz.

In this paper a broad-band coupling of IDT

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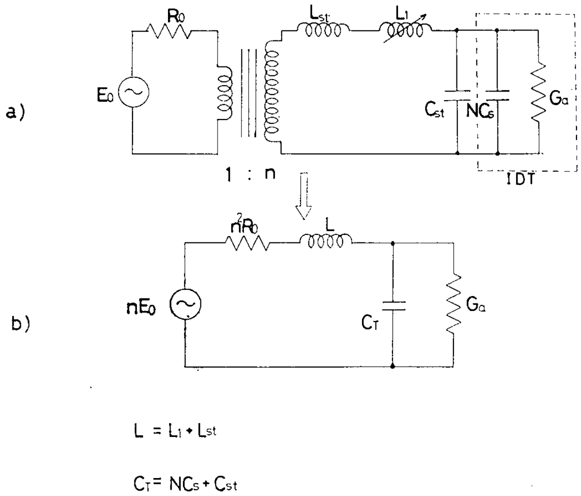


Fig. 1 a). Schematic representation of matching network interfacing between IDT and electrical source or load resistance R_0.
b). Electrical source or load is transfered into secondary circuit of transformer having turns ratio n.

is implemented by use of the matching circuit shown in Fig. 1 (a) where a step-up transformer is simplified as the ideal transformer having only one parasitic element of secondary leakage inductance. The method has the practical advantage of requiring only one transformer and one tuning inductor although the transformer has an upper frequency limit restricting the device operation to relatively low frequency.

II Disign Principle

If the condition $[(4/\pi)k^2N]^2 \ll 1$ is valid where k^2 is the SAW coupling coefficient and N is the transducer length measured in interdigital periods i. e., $2N+1$ is total electrode number, there is no distinction between the Smith's two equivalent circuits of IDT i. e., "crossed-field" model and "in-line" model[6]. The "mixed circuit" model might predict SAW-IDT more accurately but this is not the case because the end-effect manifests itself for short arrays[7]. We shall then adopt the "crossed-field" model hereafter to simplify the calculations. Figure 1 (a) depicts the broad-band matching circuit with the electrical source impedance R_0, tuning inductor L_1, a matching transformer with the turns ration n and the parasitic capacitance C_{st} which can not be neglected for IDT of short array. Radiation conductance G_a of IDT is obtained from the "crossed-field" model,

$$G_a = \omega_0 C_s k^2/\pi \left(\tan\frac{\theta}{4} \sin\frac{N\theta}{2}\right)^2 \qquad (1)$$

where $\theta = 2\pi\omega/\omega_0$ with IDT center frequency ω_0 determined from the periodic length and C_s is the static electrode capacitance of one periodic section[6]. Susceptance of IDT can be neglected for $(Nk)^2 \ll 1$. We can represent the matching transformer as an ideal step-up transformer of the turns ratio n having the series leakage inductor L_{st} in secondary circuit, which will be mentioned in the next section. The circuit can be simplified as Fig. 1 (b) where $L = L_{st} + L_1$ and $C_T = C_{st} + NC_s$. Here we neglect series resistance of electrodes against n^2R_0.

As it can be seen from eq. (1), magnitude of G_a is much smaller than ωC_T and thus the circuit is essentially series-resonant. In order to obtain broad-band matching; firstly, one can decrease the network Q factor with increase of n, while there is an optimum value of n for impedance matching between the source and IDT, and secondly, one can detune the circuit resonance frequency $\omega_1 = (LC_T)^{-1/2}$ from ω_0 to increase bandwidth. Basically, increase of bandwidth is always accompanied by a sacrifice of insertion loss; that is a trade-off relationship between the insertion loss and the bandwidth. As another approach, a parallel matching circuit with the transformer of which secondary circuit is shunted with a parallel tuning inductor and IDT may be proposed[8]. However this can not be put into practice because the transformer turns ratio n for the present purpose exceeds a realizable value.

Applying an ordinary circuit analysis, one can obtain the insertion loss defined as the ratio of power available from the electric source to the SAW power,

$$\mathrm{IL}=20\log(n^2R_0|Y|^2/4\,G_a) \qquad (2)$$

where Y

$=[(1+j\omega L/n^2R_0)\{G_a+j(\omega C_T-1/\omega L)\}-1/j\omega L]$.

Transit angle is given by

$$\phi=-\arg(Y) \qquad (3)$$

and then group delay is

$$\tau=\frac{d\phi}{d\omega}\ . \qquad (4)$$

The triple-transit suppression is obtained from the reflection coefficient of the IDT terminated on its electrical port with the matching circuit by putting $E_0=0$ in Fig. 1(b),

$$\mathrm{TTS}=-40\log|\Gamma| \qquad (5)$$

where the reflection coefficient Γ is obtained from both the models[6]. Expression of Γ is a complicated function of the angular frequency ω and parameters of the network and IDT. Though it is not minimum, Γ is written as, for $\omega\simeq\omega_0$,

$$\Gamma=\{\hat{Y}_0+1+j\pi N(\hat{Y}_0-1/12N^2+1/3)(\omega-\omega_0)/\omega_0\}^{-1} \qquad (6)$$

where $\hat{Y}_0$

$=j(1/r)\pi/(4\,Nk^2)\{1+(j\omega C_Tn^2R_0-\omega^2C_TL)^{-1}\}$ and $r=NC_s/C_T$.

All quantities of IL, ϕ, τ and TTS can be expressed as functions of the normalized frequency $x=\omega/\omega_0$, N, k^2, r, detuning factor $x_1=\omega_1/\omega_0$ and circuit Q factor $Q_1=(\omega_0C_Tn^2R_0)^{-1}$. It can be seen from eq. (1) that N should be unity at most so that G_a for a single IDT has 1.5 dB-fractional bandwidth of 50%. However, since one can not attain sufficiently low insertion loss when $N=0.5$, $N=1$ is adopted so as to realize 3 dB fractional bandwidth as well as low insertion loss.

As a design example we shall calculate IL, 3 dB bandwidth BW, the normalized frequency $x_{\min}$ which gives minimum IL, TTS and τ as functions of Q_1 with parameter of x_1 in Figs. 2 and 3 for YZ-$LiNbO_3$ with $k^2=0.046$ and $r=0.5$; the value of r is experimentally determined as it will be mentioned in the next section[9]. Here we choose $x_1>1$ to detune the network to the frequency higher than ω_0 because detuning to the lower results in a remarkable decrease in $x_{\min}$ which is close to midband frequency. In order to realize the broad-band matching, one should find the optimum condition where IL and τ are minimized while BW and TTS are maximized with the limitation $x_{\min}<1$ to avoid bulk wave exci-

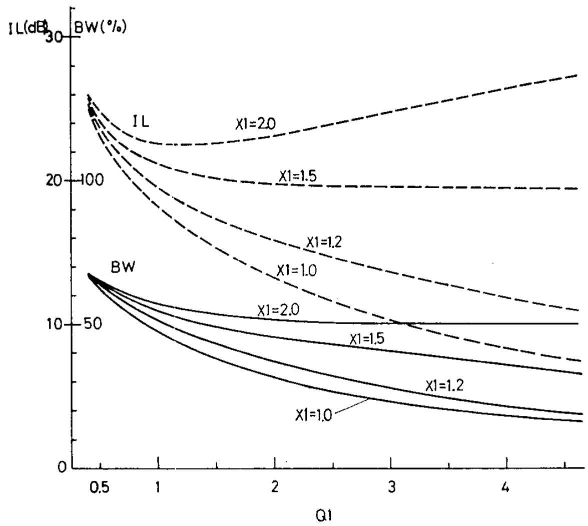


Fig. 2 Insertion loss (IL) and 3 dB fractional bandwidth (BW) for two $N=1$ IDTs as functions of $Q_1=(\omega_0C_Tn^2R_0)^{-1}$ with detuning parameter x_1.

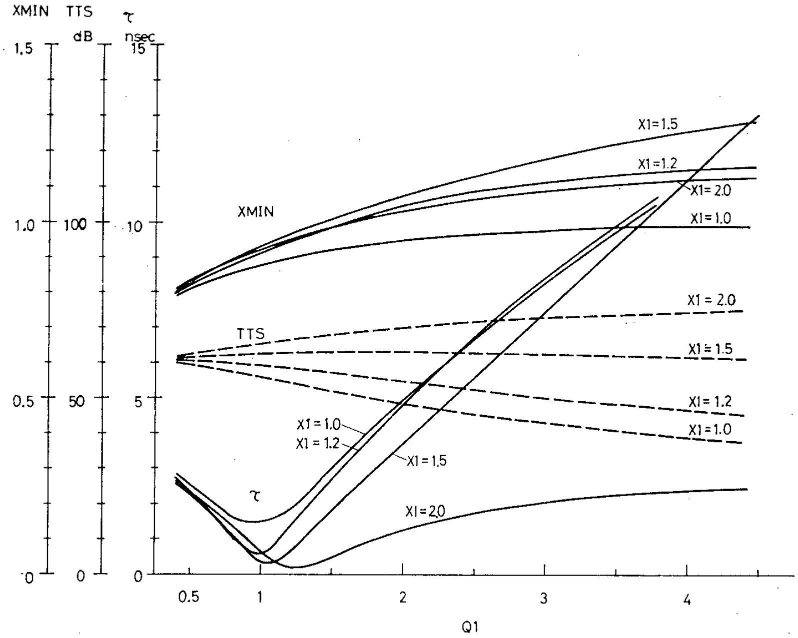


Fig. 3 Triple transit suppression (TTS), group delay (τ) and normalized frequency (x_{min}) for minimum IL as functions of Q_l with parameter x_l.

tation. We then have $Q_l=1$ and $x_l=1.2\sim2.0$. Setting $Q_l=1$ and $x_l=1.5$, results of calculations are IL=21 dB, BW=55%, TTS=62 dB, $\tau=3\,n$sec and $x_{min}=0.93$ for a single IDT. Thus the product of IL and BW is close to the maximum achievable value for YZ-$LiNbO_3$[10] and TTS is much higher than that of the ref. 3).

In the following section, using this optimum condition, we have constructed a SAW delay line with the broad-band matching circuit and a pair of IDTs having the center frequency of 35 MHz to confirm the present principle of the method.

II Experiment

A pair of IDTs spaced by 2 mm was fabricated on YZ-$LiNbO_3$, each having 3 electrodes ($N=1$), electrode width equal to space of 25 μm and aperture of 40 acoustic length (4 mm). A slant grounded electrode was made between both the IDTs. Using values of SAW velocities for free and shorted YZ-$LiNbO_3$ surfaces, the center frequency of IDT can be estimated to be 34.5 MHz[7]. The electrode capacitance $C_s=1.78$ pF is obtained from Engan's formula[11]. Total capacitance C_T and series radiation resistance R_a of the IDTs were measured to be 3.6 pF and 120 Ω, around at 35 MHz, respectively, by use of HP-4815 A vector impedance meter. Thus the ratio r is 0.49 and R_a agrees reasonably with the theoretical value of maximum radiation resistance 142 Ω calculated from $R_a\simeq G_a/(\omega_0 NC_s)^2$, 6).

The transformer turns ratio n which gives $Q_l=1$ amounts to 5 by assuming $R_0=50\ \Omega$ for the conventional system. Wide-band transformers having the center frequency of 35 MHz and $n=4\sim7$ were made of Philips 4 C 4 troidal cores (7 mm in diameter and $\mu\simeq100$). The primary coil was designed to have sufficiently high impedance compared to R_0, with a typical example of 5.3 turns and 0.59 μH. This was wounded over the whole circumference of the core with equal space. The seconary coil of which impedance must be sufficiently greater than R_a was typically 28 turns and 13 μH. It was wounded on the primary coil so as to minimize the leakage inductance L_{st}, being 1.6 μH. The secondary coil gives impedance of about 3 kΩ at 35 MHz. Actually, however, the secondary circuit was parallel resonant at 36 MHz with the resultant impedance of 7.6 kΩ, indicating inclusion of the shunting parasitic capacitance of 1.5 pF. By parallel connecting two transformers on both the secondary terminals, 1.5 dB-fractional bandwidth of each transformer was found to be 90% centered at 35 MHz, leading to the conclusion that one may regard it as an ideal transformer except the secondary leakage inductance L_{st}. Accordingly the circuit analysis and the method of design in the preceeding section are valid. As for the tuning inductor which gives $x_1=1.5$, inductance L is calculated to be 2.43 μH and thus L_1 is 0.83 μH by subtracting L_{st} from L. These inductors, transformers and YZ-$LiNbO_3$ plate were mounted on a printed board.

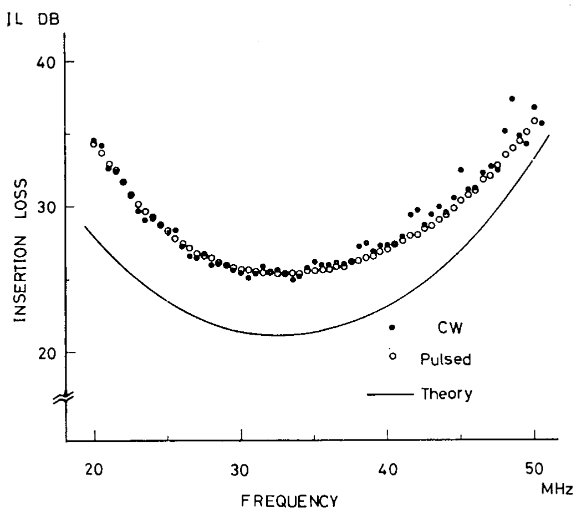


Fig. 4 Measured and calculated insertion loss (IL) against frequency; ● (dot) represents IL for CW input signal and ○ (white circle), for pulsed input.

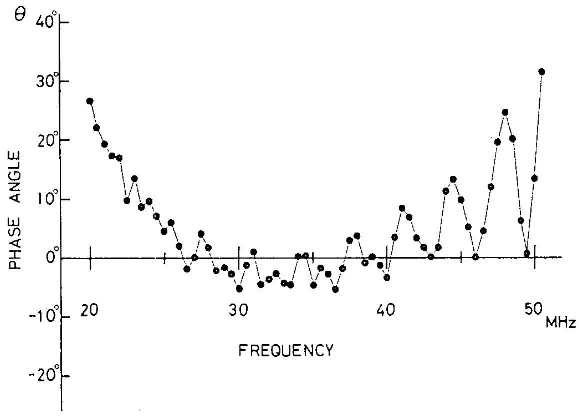


Fig. 5 Measured phase angle against frequency (CW input signal).

IV Results and Discussions

By use of HP-8401 network analyser, insertion loss and transit angle were measured as functions of frequency as shown in Figs. 4 and 5, respectively. In Fig. 4 dots show IL obtained from CW input signal and white circles, from pulsed signal. It can be seen from Fig. 4 that 3 dB pass band ranges from 15.7 to 41 MHz and the frequency which gives minimum IL of 24.8 dB is 33 MHz, indicating $x_{min}=0.94$ and BW=46.5%. The curve calculated from eq. (2) agrees fairly well with the experimental data, which show higher IL and smaller BW than the theoretical predictions.

Figure 6 shows a typical output waveform with 0.2 μsec pulse-modulated input signal. Arrow 1 indicates the feed-through signal. The main time lobe lags by 0.6 μsec as shown by arrow 3, confirming the SAW velocity of 3.49×10^3 m/sec on free surface of YZ-$LiNbO_3$. Triple transit echo, which is too small to be identified in Fig. 6, appeared 1.7 μsec after the input signal. TTS level is plotted against frequency, having the minimum of about 61 dB

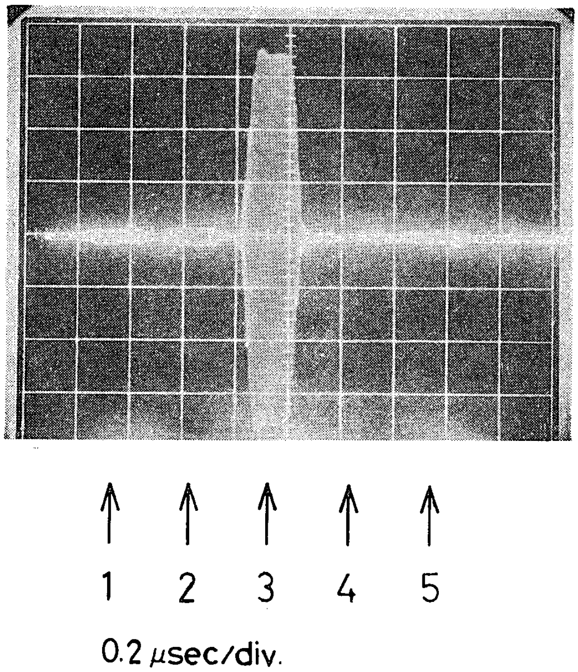


Fig. 6 Waveform at receiving IDT for RF pulse of 0.2 μ sec duration. Arrow 3 indicates SAW output and others, spurious signals.

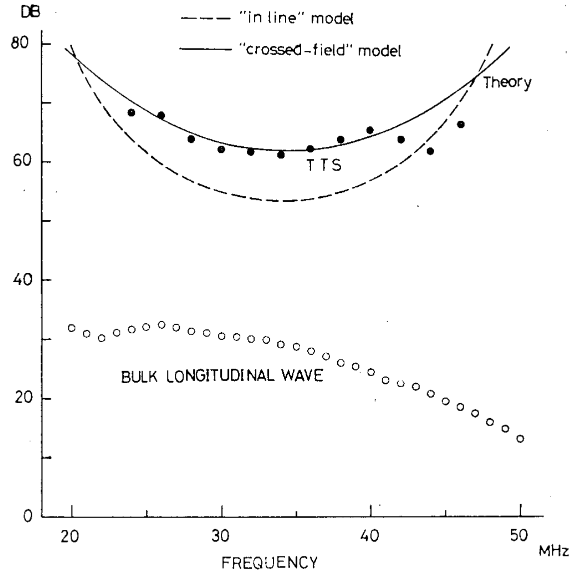


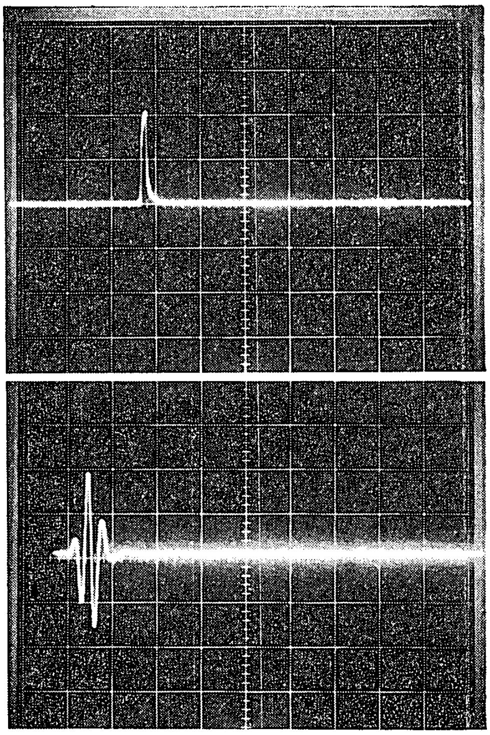
Fig. 7 Measured and calculated triple transit suppression (TTS) as a function of frequency and measured power ratio of SAW to bulk longitudinal wave vs. frequency (○ white circles).

at the midband frequency of 33 MHz (Fig. 7). The spurious signal indicated by arrow 2 in Fig. 6 has the delay time of 0.28 μ sec, predicting velocity of 7.1×10^3 m/sec which is close to that of bulk longitudinal wave of 6.9×10^3 m/sec and thus this signal is identified with the longitudinal wave diffracted into the bulk of *YZ*-$LiNbO_3$,[12]. In Fig. 7 white circles show the ratio of the SAW power to the bulk longitudinal wave power against frequency. The spurious signal increases with increase of frequency, amounting to −23 dB of the SAW signal at the upper edge of 3 dB pass band. On the other hand the triple transit echo is less than −30 dB of the SAW signal over the 3 dB pass band and agrees well with the curve calculated from the "crossed-field" model (the solid curve in Fig. 7) but disagrees with that from the "in-line" model (the dotted curve), contrary to the ref. 3). The ripple of phase angle having a period of 3.5 MHz in the frequency axis can be seen in Fig. 5 and explained as the interference between the SAW signal and the bulk longitudinal wave if the difference of delay time between them is taken into account. Furthermore the scatter of CW output signal (dots) in Fig. 4 can be explained in the same way. If this ripple i.e., bulk longitudinal wave can be removed, one can attain BW of 50%, an agreement with the theoretical prediction of 55%. Since the spurious indicated by arrow 5 has the delay time more than two times larger than that of the SAW signal, we can identify it with the slow shear wave reflected at the base of the sample[12~14]. This signal also increased with increase of frequency, leading to −36 dB of the SAW signal. The spurious signal pointed by arrow 4 can be considered as the reflected fast shear wave, of which power is much smaller than the others (Fig. 6). Though the fast shear mode is piezoelectrically inactive

SH-mode in YZ-$LiNbO_3$, deviation from coincidence between sagittal plane and YZ-$LiNbO_3$ plane due to finite aperture of IDT makes this mode piezoelectrically active. One can suppress the reflected bulk waves by making the base of the sample rough and/or by use of various absorbing compounds. However the spurious signal due to bulk longitudinal wave can be suppressed only by separating transmitting and receiving IDTs sufficiently and by use of a multistrip coupler[15,16]. Milsom et. al. reported that the loss due to bulk shear wave is substantial for short arrays, contrary to the present experimental results[12]. In order to clarify this discrepancy, it is necessary to examine the angular distribution of bulk longitudinal wave power for $N=1$ and effect of source impedance on it. Slight misorientation and imperfections of the crystal also may have accounted for this as well as launching the fast shear wave.

a)

b)



0.1 μsec/div.

Fig. 8 Video pulse time respones of SAW delay line with two $N=1$ IDTs. a). input waveform and b). output waveform.

In Fig. 8 we can see the shape of output signal for input in the form of a video pulse with 20 *n*sec duration. The duration of the output including significant time sidelobes is 70 nsec, suggesting that the present method can be applied to recirculating memory stores.

V Conclusions

By use of only one step-up transformer to decrease the matching network Q and one tuning inductor to detune the network from the IDT center frequency, we have proposed a broad-band impedance matching between IDT and conventional electrical source or load resistance of 50 Ω. This method predicts octave bandwidth operation with relatively low insertion loss and very high TTS up to the frequency limit of the transformer. An experimental verification has been made, exhibiting the results of insertion loss of 25 dB, fractional bandwidth of 46.5% and TTS of 61 dB at the midband of 33 MHz, in agreement with the theoretical calculations but for the bandwidth. Degradation of the bandwidth against the predicted value of 55% is substantially due to bulk longitudinal wave excitation, which differs from the existing literatures.

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